How Superheterodyne Circuits Are Made To Reject Unwanted Signals

LESSON ND-24

SPRAYBERRY ACADEMY of RADIO

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An important part of a radio wave which you have not studied up to now is the so-called *side-band* frequencies which accompany the radio wave. These side band frequencies expand the carrier to other frequencies which are plus and minus that of the carrier. An example will make this clear. Suppose a radio station has a carrier frequency assigned to it by the Federal Communications Commission equal to 700,000 cycles or 700 KC. It is the custom of the Commission to assign carrier frequencies to other stations on adjacent carrier frequencies. Thus, above 700 KC the nearest adjacent carrier frequency will be 710 KC, and below 700 KC the nearest adjacent carrier frequency will be 690 KC. Thus, it is seen that radio stations are spaced 10,000 cycles or 10 KC apart in the broadcast spectrum. From 540 to 1600 KC there is a channel (space for carrier frequency) every 10 KC. Some stations use a certain channel exclusively and other stations share the same channel with still other stations—all under the supervision and rules of the Commission. Those stations which share the same channel are so regulated as to power and time of operation that there will be a minimum of interference between stations. It is in this way that so many stations can operate in the broadcast band.

Returning to the example of the 700 KC carrier frequency, its property right in the frequency spectrum is 700−5 or 695 KC and 700 +5 or 705 KC. If its signals should extend to 694 or to 706 KC it would interfere with the 690 and 710 KC channels. In other words each station is permitted to radiate signals up to 5000 cycles or 5 KC each side of its assigned carrier frequency. This gives every radio station a width of 10 KC which it must not exceed.

When a radio carrier wave is modulated, it is said to be accompanied by two other frequencies. If, for purpose of illustration, a constant modulating frequency of 3000 cycles is assumed and the carrier is 700 KC as mentioned, the other frequencies accompanying the carrier will be equal to 700−3 or 697 KC for one frequency and 700+3 or 703 KC for the other frequency. These two additional frequencies are solely the result of mixing the modulating frequencies with the carrier frequency to form a complex radiated wave form. The two extra frequencies are often called *side band* frequencies. The plus frequency, as in the foregoing case of 703 KC, is called the upper side band frequency, and the minus frequency, as in the case of the 697 KC, is called the lower side band frequency. In this case, a single frequency of 3000 cycles has been

used as an example to illustrate the principle involved. Actually, as you know, in practice radio stations do not impress upon their carriers a single sound frequency as in the case of the 3000 cycle example. Instead, audio frequencies from about 40 to 5000 cycles are being continuously impressed on the carrier as this is about the usual AF frequency range of the average broadcast station. Thus, if 40 cycles is accepted as a minimum and 5000 cycles as a maximum and if you keep in mind that from 40 to 5000 cycles are continuously impressed on the carrier, it is easy to see that for a 700 KC carrier it is going to expand to 700,040 cycles as a minimum and up to 705,000 cycles as a maximum. Too, this process is going to be a continuous variation because the very nature of sound frequencies is to vary between extreme limits.

The purpose of this brief discussion of the carrier frequency and its accompanying side band frequencies is to bring out the point that tuned circuits (including RF and IF) must not be so arranged as to tune to the carrier frequency only. If this were done, the modulating frequencies would be tuned out entirely, and there would be no AF to operate the speaker after detection. Instead, when a tuned circuit is adjusted to accept a 700 KC carrier, for example, it must be so broadly tuned as to also accept the lower and upper side band frequencies. The ideal tuned circuit for a 700 KC carrier would accept all frequencies down to 695 KC and all frequencies up to 705 KC. Beyond these extreme limits it would sharply cut off and reject all other frequencies so that there would not be adjacent channel interference from the 690 and 710 KC and other nearby carriers. However, this ideal is never attained and so compromises must be made with everything involved in order to approach as near as possible to the ideal. You will learn about these factors in this lesson. The ability of a tuned circuit to reject unwanted signals is known as *selectivity* ability and this, too, will be explained later on.

Thus, from this discussion you should understand that tuned circuits must be designed and adjusted to *accept a relatively wide band of frequencies and not just the carrier alone.*

The subjects of modulation, side band frequencies, selectivity, etc., will be taken up again further on in this lesson and in other lessons to follow. For the time being your attention is directed to the following discussion of tuned circuits. With the qualifying factors of these properly understood, you will be in a better position to appreciate the later discussion on modulated carriers, their side band frequencies and response curves of tuned circuits.

A number of very important subjects are to be studied in this lesson such as tube characteristics, tuned circuits, coupled circuits and filter systems. As you proceed with this study, you will find that the value, position and other properties will affect all others, and often a compromise must be sought which will give the best results for the problem under consideration. Many times one value for a

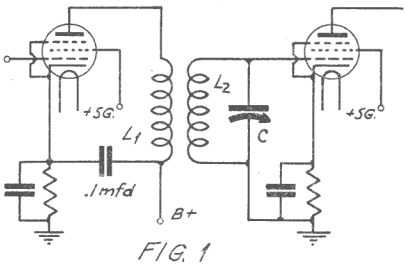


FIG. 1

given part will be best from one point of view but undesirable from another. In these situations a value must be selected, not from calculation, but from judgment of how much of the desirable property may be used at the expense of using part of an undesirable property.

In Fig. 1 is shown a standard RF amplifier using an RF pentode tube with a tuned secondary transformer (L2). Considering first the gain of the tube, you will recall that this depends on the amplification factor of the tube and the ratio of its plate circuit load resistance or impedance to its AC plate resistance. The amplification factor of the tube is fixed by its manufacturer, and after a tube of the proper characteristic has been selected, a plate load suitable for the performance must be found. All of the influences tending to affect coil L1 in Fig. 1 serve to make up its impedance, including its own self-inductance. However, this latter property of the coil is small in comparison to other factors in making up its total impedance. The term impedance is used here instead of reactance, as *the effect of the coil is that of inductive reactance and DC resistance of the wire winding.*

Since it is already known that when two coils are magnetically coupled there will be *mutual inductance between them,* due to the influence of one on the other, it is necessary to see how the secondary (L2 of Fig. 1) affects the primary impedance (L1). When the value of the complete primary impedance is known, it is possible to calculate the gain of the stage by making use of formulas which have already been given in other lessons. Besides having a very marked effect on the primary, the secondary has *another purpose*—that of forming the inductance (L2) of a tuned circuit. *That is its most important function* and serves to most accurately determine the requirements for its design. Therefore it is well to first investigate the design of the secondary and then the primary.

The inductance value of the secondary winding is not a fixed value, but it depends on the frequency and the capacity of the tuning condenser. The formula relating to the capacity, inductance and frequency shows that as long as the product of L and C remains unchanged, the frequency will remain unchanged.

Thus the resonant frequency formula, as you have learned, is equal to:

$$(1) \qquad F = \sqrt{\frac{25{,}330}{LC}}$$

Where—L = microhenries
C = microfarads
F = Kilocycles

As you will learn later on, it is also desirable to know the product of LC. In formula form this is equal to:

$$(2) \qquad LC = \frac{25{,}330}{F^2}$$

As in formula 1, the units are in microhenries, microfarads and kilocycles.

Now to return to the secondary inductance value of coil L2 in Fig. 1, suppose a frequency value of 1000 KC is selected to illustrate the principles involved. The product of L and C for the tuned circuit consisting of L2-C in Fig. 1 is equal to:

$$LC = \frac{25{,}330}{1000^2}$$

$$= \frac{25{,}330}{1{,}000{,}000}$$

$$= .02533$$

Thus, at 1000 KC the value of the coil L, in microhenries is equal to $\frac{.02533}{C}$ and the value of C in microfarads is equal to $\frac{.02533}{L}$.

Next the question arises of how many combinations of L and C can equal .02533 when multiplied. The answer is an indefinite number. A few examples are included in the following tabulation to illustrate just what is meant by this statement.

CAPACITY (Microfarads)	INDUCTANCE (Microhenries)	PRODUCT of L×C
10	.002533	.02533
1	.02533	.02533
.02533	1	.02533
.002533	10	.02533
.002	12.65	.02533
.001	25.3	.02533
.00035	72.25	.02533
.0002533	100	.02533
.00005	506.5	.02533
.000001	25,330	.02533

It should be noted that all of these combinations will tune to the same frequency. Why, then, you may ask, is one set of values chosen instead of another? The answer is that for a fixed frequency there is only one thing which determines the proper or most efficient combination of L and C values. This is the *coil efficiency or the ratio of its inductance to its radio frequency resistance.* For convenience, a symbol for this ratio

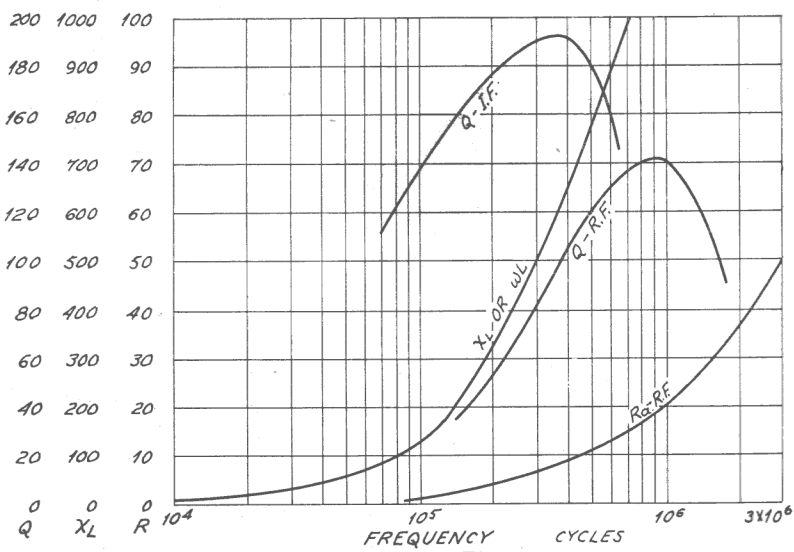


FIG. 2

is used. It is derived from $\frac{2\pi fL}{Ra}$, which is called the Q factor of a coil. Ra, in this instance, is equal to the *apparent radio frequency resistance* of the secondary or tuned winding.

The distributed capacity in farads between wire turns and across the ends of a coil together with its inductance serves to determine the radio frequency resistance of the wire winding to a great degree. The DC resistance of the wire winding must also be considered.

It should be noted that the *apparent resistance* (Ra) of a coil of wire at RF is due to the so-called *skin or surface effect* of a wire winding. It is known that *high frequencies* do not travel inside a wire to any large degree—rather they are known to travel on the surface of the wire — hence, the more wire surface, the less the RF resistance. More about this later.

You can easily visualize how the condenser effect between wire turns would affect the high frequency currents traveling on the surface of the wire. Obviously it is impractical to undertake to compute this Ra value by determining the new distribution of electrons on the wire surface. Since the capacity between wire turns has an increasing effect on the distribution of the RF current, it is related to Ra and can be used in finding its value. The way in which these factors—that is, L, R and C of a coil—are related to Ra, the apparent RF resistance is as follows:

$$Ra = \frac{R}{(1-\omega^2 LC)^2}$$

Where:

Ra = The apparent RF resistance.
R = The DC resistance of the wire winding.
ω = $2\pi f$ or $6.28 \times f$.
L = The inductance in henries.
C = The distributed capacity across the coil turns in farads.

The symbol ω is the Greek letter Omega, and note it is equal to $6.28 \times f$. It will be used again later on so be sure to remember its significance.

An average coil of 200 microhenries has a distributed capacity of 6.34 mmfd. It will also have an average resistance of 12 ohms. Both the distributed capacity and RF resistance of a tuned coil are difficult to calculate or measure. For a given type coil a series of measurements are usually made in the laboratory and average values are arrived at which may be used for all coils of that same general type.

Actually, the true RF resistance value is exceedingly hard to obtain. Its value is either assumed in the design of coils (based on experience) or it is measured by expensive and complicated bridge circuits.

You have learned in previous lessons that it is desirable to have an impedance which is as high as is possible for the control grid of a tube. You will remember that the selection of the L and C values determines the shunt impedance of any tuned circuit at resonance, provided other factors do not change. Thus, a circuit that has high inductance also has a high shunt impedance. You also learned from a previous lesson that for

a constant value of inductance, the lower the capacity the higher will be the *shunt impedance* of the tuned circuit. It may also be shown that for one frequency, this impedance increases and as the capacity decreases.

This being so, you may wonder why it is not desirable to use such values as 25,330 microhenries and .000001 microfarads (see the preceding table) for the tuned circuit of Fig. 1. These values represent a very high inductive impedance and a very low capacity. The answer is that the efficiency of a resonant circuit or its ratio of reactance squared to its resistance ratio $\frac{\omega^2 L^2}{Ra}$ *determines the sharpness of its tuning* (its selective ability). It is possible to increase the inductance and decrease the capacity until a point is reached where the *radio frequency resistance increases faster than the inductance of the coil,* making $\frac{\omega^2 L^2}{Ra}$ increase less rapidly and making $\frac{\omega L}{Ra}$ or Q reduce in value.

Increasing the inductance is an easy matter, but in order to have *as little radio frequency resistance as possible* and to build the inductance in a fairly small space, the wire of which it is wound must be small and of considerable *length.* This greatly increases its radio frequency resistance which, as you have already studied, depends on the wire surface area, DC resistance, dielectric of the coil form on which it is wound and its insulation.

Figure 2 shows a graph of the manner in which the Ra resistance of a coil of wire increases with frequency, due to the tendency of the RF to travel on the skin or surface of the wire. In this graph Ra is compared to the inductive reactance and to the average practical Q value of the circuit. Above 300 to 400 microhenries for radio frequency in the broadcast band (540 to 1600 KC) an effective coil cannot be wound in a reasonable space. To get an effective coil of this inductance, often it is wound *with stranded wire* (which gives more surface area) and with a large diameter as compared to its length. With a diameter of approximately 2½ times its length, a coil will have maximum inductance for the length of wire used in a single layer coil. This is not always convenient, and it is not observed rigidly in coil manufacture.

In the study of inductance, mention was made of the fact that between each turn of wire there existed a capacity. This, of course, depends on the wire shape, size and spacing. Coil capacity adds to any capacity used to tune it and the former must be kept at a minimum to avoid too large a total capacity. For a single layer coil of the conventional type, the coil capacity is proportional to the number of turns.

It is these two factors, the radio frequency resistance and the inherent coil capacity, that limit the size of the inductance which would be practical. Now it must be remembered that all coils have these properties and the selection or design of a coil remains with the design engineer. Just how much radio frequency resistance or capacity can be tolerated in view of the advantages of high inductance is his problem. For this reason,

all radio receivers do not employ the same size coils or tuning condensers. Perhaps for advantages of mechanical construction, shielding, spacing of parts or other electrical requirements, various values are used.

GAIN OF THE AMPLIFIER

By this time you may think that the discussion has completely wandered away from the original problem of designing the primary of an RF transformer, but all of the foregoing has a definite bearing on the primary.

There being no iron or other magnetic metal core materials in most RF coils, the magnetic fields are spread out over a considerable area. For this reason non-magnetic metal shielding must be used to prevent the field of a coil from influencing other coils or windings. The shield should be as far away from the surface of the coil at every point as the coil diameter so that it will not create too much loss in efficiency of the coil through current induced into it. Any non-magnetic shielding material such as copper, zinc or aluminum will *decrease the inductance of any coil* over which it is placed. This is due to the fact that it will act as a shorted-turn secondary for any leakage magnetism reaching it. For proper shielding, this decrease will only be a small percentage. Magnetic material such as iron or sheet steel will increase the inductance of a coil when used as shielding because of its magnetic properties in serving as an outside core. Its effect is to lower the reluctance of the magnetic path, and, since magnetism may be formed more easily, the inductance becomes higher by a small percentage.

Because of the fact that there is no restricted magnetic path in an RF air core transformer many things which apply to iron core transformers do not apply to the former. The coupling of one winding to another in an iron core transformer is always considered as complete or maximum, whereas the spacing between high frequency coils, their positions with respect to one another and their relative sizes determine the amount of coupling of air core transformers. You can see that since there must be a great deal of magnetic leakage, one coil of an RF transformer will not necessarily be influenced by all of the magnetic lines of force of another which is coupled to it. It is possible with an RF air core transformer to so space one winding that it cannot be influenced by another adjacent to it. In this case, the coupling is essentially zero.

Coupling may be regarded on a percentage basis, calling it 100% when maximum and 0% when zero, but it is mathematically more convenient to measure coupling in terms of a coefficient or factor. In this way, maximum coupling is considered as one and zero coupling as zero. If the coils are 50% coupled, the coefficient of coupling is 0.5. Quite naturally, maximum coupling is obtained when the windings of one coil are close to and parallel with those of the other.

For a time it was thought that the greatest amount of energy would be transferred between two tuned coils if they were coupled a

maximum amount and so the two windings were wound simultaneously on a single coil form. Between each secondary turn of the finished coil was a turn of the primary. Later a method of placing one winding over the other was thought to be satisfactory.

Now in the study of inductance it was learned that when current changes its intensity in a coil, a voltage is induced which accounts for the reactance of the coil. In addition to this, you know that every current is accompanied by a magnetic field. When two coils are closely coupled, their fields will interlock and one coil will influence the reactance of the other through their mutual inductance.

The opposition to current flow in the secondary at resonance, due to the primary current *has the effect of resistance in series with the secondary circuit.* Resistance either in shunt or in series with a tuned circuit *will lower its efficiency,* but for simplicity in dealing with such circuits, always deal with the resistance *as though it were in series with the resonant circuit.* This is done simply to reduce all considerations to a common basis. Unlike the ordinary DC resistance, due to the metal of which the conductor is made, the *RF resistance* increases as the current flow increases. This is because the radio frequency current redistributes itself in the cross section of the wire, using only the conducting path of the wire near the surface. This action is much the same as though the diameter of the wire were decreased as the frequency applied to it is increased. Added to this effect is the fact that with maximum current a maximum magnetic field exists about the tuned coil. This field will oppose the original field which produces the voltage in the coil. From all of these effects it will be seen that the resonant current in the secondary circuit is going to be reduced considerably.

With these facts in mind, refer to the curve in Fig. 3 of the peak voltage of an *unmodulated* radio frequency wave as it appears across the tuned circuit for a range of frequencies through resonance. At F1, somewhat below resonance, the tuned circuit carries practically no current, making the two coils (L1 and L2 of Fig. 1) have essentially *no mutual effect* upon each other. The reactance of the plate coil L1 is quite low; in fact, the amount present is due only to its own self-inductance.

As resonance is approached, current through and voltage across the tuned secondary increases. It is maximum at resonance and decreases above resonance in practically the same manner that it increases up to resonance. Now see what effect this has on the primary or L1 of Fig. 1.

When one circuit is *coupled* to

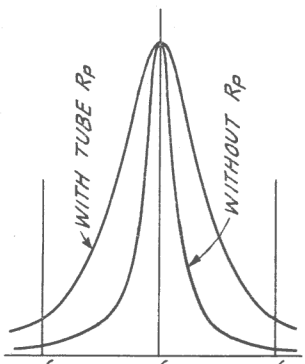


FIG. 3

another, not only is it possible to transfer energy from one to the other, but one will entirely change the characteristics of the other. To clarify this point, first consider an iron core power transformer as in an ordinary radio receiver. With the secondary circuit open, and an AC ammeter in series with the primary, the current measured will be exceedingly small. It will be only enough to account for the slight heating losses of the transformer. Perhaps no more than 50 milliamperes of current will flow. This, of course, would indicate that the reactance of the primary would be

$$\frac{110}{.05} = 2{,}200 \text{ ohms.}$$

On connecting the secondary to a load, or, in fact on loading the transformer with several secondaries connected to various devices, the primary may be made to carry up to 500 ma. or .5 ampere with safety. Without doing anything to the primary, its reactance has changed from 2,200 to 220 ohms. If a very high resistance is connected in series with each secondary, the load is reduced and this has accomplished the same effect as placing a resistance in series with the primary.

Somewhat the same conditions take place with RF transformers. The reactance of their primaries depends vitally on the electrical nature of their coupled secondaries. For example, in a circuit like Fig. 1, the primary circuit is affected exactly as though an *impedance* has been placed in the primary.

The *shunt impedance* of a tuned circuit is $\frac{\omega^2 L^2}{R}$, and, since there is mutual inductance between the coils, L may be replaced by M (M is the symbol for *mutual inductance* as explained in a former lesson) in expressing the impedance appearing in the primary due to the action and relation of the secondary to it.

Thus, $\frac{\omega^2 M^2}{Zs}$ expresses the value of *impedance coupled into the primary* (from the secondary) and acting as though it were in series with it. Zs in this case equals the impedance of the secondary. Not only does the mutual inductance act in both circuits, but each individual inductance acts in its own circuit as well. Therefore, the complete expression for the primary impedance is:

$$Zp = Zo + \frac{(\omega M)^2}{Zs}$$

Where:
Zp = Primary impedance.
Zo = Self-inductive reactance of the primary.
Zs = Series resistance of the tuned secondary.

You will find further that M is equal to the coefficient of coupling, or K times $\sqrt{Lp \times Ls}$.
Where:
Lp = Primary inductance.
Ls = Secondary inductance.
K = 1.

As will be pointed out later, the effective impedance of the primary has a very marked effect on the impedance of the secondary. Two examples will be used in the following to show how different values of inductance in the primary will create different primary impedance values (Zp). This Zp can be effectively transferred or coupled into the secondary circuit as will be pointed out later.

As pointed out in the foregoing, the primary impedance is equal to:

$$Zp=Zo+\frac{(6.28\times f\times K\times\sqrt{Lp\times Ls})^2}{Zs}$$

If values are assigned to Zo and L and Zs, the Zp formula may be used to show how the primary impedance is obtained. For the first example, let Ls equal 200 microhenries, Lp equal 50 microhenries, Zs equal 8 ohms and f equal 1000 KC or 1,000,000 cycles. For the second example, all values will remain the same with the exception of Lp which will be 100 microhenries instead of 50 microhenries. For these values note that 200 microhenries is equal to $2\times^{-4}$ henries; 50 microhenries is equal to 5×10^{-5} henries and 1000 KC is equal to 10^6 cycles. Because the coefficient of coupling K, in this case, is assumed to be equal to 1, it may be omitted from the examples to follow. In the example where Lp equals 50 microhenries, the primary impedance equals:

$$Zp=6.28\times10^6\times5\times10^{-5}+\frac{(6.28\times10^6\times\sqrt{5\times10^{-5}\times2\times10^{-4}})^2}{8}$$

When a problem is set up in this form, the first step is to find the equivalent value of the terms outside the parentheses. This step follows:

$$\begin{aligned}&6.28\times10^6\times5\times10^{-5}\\&=6.28\times5\times10\\&=6.28\times50\\&=314\end{aligned}$$

The next step follows:

$$Zp=314+\frac{(6.28\times10^6\times\sqrt{5\times10^{-5}\times2\times10^{-4}})^2}{8}$$

The value within the square root sign is equal to:

$$\begin{aligned}&\sqrt{5\times2\times10^{-9}}\\&=\sqrt{10\times10^{-9}}\\&=\sqrt{10^{-8}}\\&=10^{-4}\\&=.0001\end{aligned}$$

The complete problem should now read:

$$\begin{aligned}Zp&=314+\frac{(6.28\times10^6\times10^{-4})^2}{8}\\&=314+\frac{(6.28\times10^2)^2}{8}\\&=314+\frac{(628)^2}{8}\\&=314+\frac{394{,}384}{8}\\&=314+49{,}289.8\\&=49{,}603+\text{ ohms.}\end{aligned}$$

Thus it is seen that under the conditions as stated the primary with an Lp of 50 microhenries has an impedance (Zp) of 49,603 ohms. Now if Lp is increased to 100 microhenries with Ls remaining at 200 microhenries, calculation according to the foregoing method will show that the primary impedance increases to 64,695 ohms. All of the foregoing assumes, of course, that the coupling is maximum between the primary and secondary. Later in this lesson it will be shown that less than maximum coupling is used in practice and that minimum and maximum inductance values are, at times, used for Lp.

If it were not necessary to consider another factor, the foregoing would provide all the information necessary to design RF transformers. Since the secondary has such a major effect on the impedance of the primary, *it will be seen*

that the primary must also have an effect on the impedance of the secondary because there is a mutual relation between them.

As has been stated, if the secondary is open, out of resonance or does not draw energy from the primary for some other reason, the primary impedance will drop down to practically the same value as the inductive reactance of the winding. Looking at this situation the other way around, the primary cannot materially affect the tuned secondary if the former is not loaded in some way. As you know, being in series with the plate circuit of the RF tube, the primary winding is loaded with the AC plate resistance of the tube. This means that any mutual inductance between windings must affect the secondary as much as the primary. Although this mutual inductance is effectively in shunt (parallel) with the secondary, *it has the effect of a series resistance on it at resonance* because it reduces the current. *Therefore it is dealt with as though it were a resistance in series with the resonant circuit.* The amount or value of this series resistance depends on the AC plate resistance of the tube connected to the primary, the self-inductance of the secondary and the coupling betweeen the primary and secondary.

It is obvious that if the coils are coupled too tightly in order that the total primary impedance may be as high as possible this will automatically reduce the selectivity of the secondary, due to the increased mutual resistance. When the coupling is tight or near maximum, the mutual inductance affects the secondary so much *that its selectivity is somewhat reduced due to increased series resistance.* When the coupling is too low, the effect of the mutual inductance, which is small in this case, will not be appreciable on the primary. The reason for this is the tube is working into such a small load that its gain will be low. *Thus, a compromise must be decided upon for the correct relation between selectivity and gain, as one must be sacrificed for the other.*

For this reason there has been a wide variety of transformer designs used in receivers. Some manufacturers have used a large inductance as high as 2 to 5 millihenries for the RF transformer primary loosely coupled to the secondary, and others have used a very small inductance, 25 to 100 microhenries, coupled very closely to the secondary. This is one point in which the design engineer must use his own judgment. The total gain and selectivity ratio wanted is usually what controls the exact design of the RF coil.

You will realize at once that the RF transformer to be used in a given tuned circuit will depend on the tubes used for amplification because different Rp values affect tuned circuits differently through mutual inductance. For triodes with inherently low Rp values, the coupling between the primary and secondary must be quite low, as the Rp of the tube loads the tuned circuit quite heavily through mutual inductance. *For RF screen grid tubes, the Rp is much higher and hence much closer coupling may be used.* Even so, only a small amount of the total amplifying power of the tube can be util-

ized. For RF pentodes having even higher Rp values, large primaries with still closer coupling may be used. In fact, tube circuits have never been developed which can take full advantage of the RF pentode tube.

It is necessary now to investigate just how the primary affects the secondary series resistance and why it is desirable to have the RF tube AC plate resistance as high (in series with the primary) as possible.

You have previously learned that the shunt impedance of a tuned circuit, as in Fig. 1, is equal to $\frac{\omega^2 L^2}{Ra}$ where Ra is the RF series resistance of the circuit at resonance. Ra, as you have learned, is principally due to the high frequency resistance of the wire winding of the coil.

But due to the *effect of mutual inductance* between the primary and secondary windings, there will be an unusual *effect in the tuned secondary circuit.* What actually happens is that *a portion of the primary impedance is, in effect transferred to the secondary circuit. This is a new impedance,* and is in addition to the Ra resistance previously mentioned. This new impedance may be called Rm, because it is entirely due to mutual inductance between the two windings.

Remember that when a tuned circuit is at resonance, it acts just like a resistance. Now consider the tuned circuit of Fig. 1 as being at resonance. Then *imagine* that the Rm or *mutual resistance* is transferred to the secondary circuit from the primary circuit. This new resistance (imaginary) is actually in parallel with the tuned circuit—you may think of it as being connected directly across the tuned secondary winding.

The mutual resistance is designated as Rm. The other impedance (that present without the effect of the primary) is expressed as $\frac{\omega^2 L^2}{Ra}$ or Z which is the impedance of the tuned secondary alone (without the effect of the primary).

From this it is seen that there are three impedances. These are:

Rm = The mutual impedance (due to the effect of the primary.

Ra = The high frequency resistance of the secondary winding.

$\frac{\omega^2 L^2}{Ra}$ = The total impedance of the tuned circuit (without the effect of Rm).

When the effect of Rm is added to the total secondary impedance, it will be found that Rm and $\frac{\omega^2 L^2}{Ra}$ are in parallel.

Now, if two resistances (or impedances) of a tuned circuit are in parallel, they may be expressed just as for any parallel circuit, because of resonance the tuned circuit acts just like a resistance—the reason for this, you will remember, is because the capacitive and inductive reactances cancel out, leaving only the resistance of the circuit.

If desired, the effect of Rm and Z may be treated as a parallel circuit, to finally obtain the value of the overall impedance of the tuned secondary circuit. However, the problem is made much simpler by

resolving Rm as an effective resistance *in series with Ra.* This makes the whole problem much simpler and leaves only a simple series circuit to work with.

Due to the fact that Rm may be resolved into a resistance which will be in series with Ra, it is usual to designate the *new series value* as Rs. Once this is done, you may forget about Rm and consider Rs in its place.

$$Rs = \frac{\omega^2 L^2}{Rm}$$

When this has been done, Rs, in effect, is in *series* with Ra which, you will remember, is the high frequency resistance of the wire of the coil. Now when two resistances are in series, their total value is simply their sum. Therefore:

$$Ra+Rs=\frac{\omega^2 L^2}{Z} + \frac{\omega^2 L^2}{Rm}$$

This may be further simplified to:

$$Ra+Rs=\frac{\omega^2 L^2 Rm+\omega^2 L^2 Z}{Z\ Rm}$$

And after the term $\omega^2 L^2$ has been factored out of the equation, it leaves a result of:

$$Ra+Rs=\frac{\omega^2 L^2\ (Z+Rm)}{Z\ Rm}$$

If the value of Rs alone is desired, it is equal to:

$$Rs=\frac{\omega^2 L^2\ (Z+Rm)}{Z\ Rm} - Ra$$

To further illustrate how Rm (from the primary) affects the tuned seconary, values of Rm may be assumed which will permit the value of Ra+Rs to be obtained—this being formed in the tuned secondary circuit.

The following table shows four typical solutions of this problem where four *common values* of Rm are assumed. In these typical problems, the following values have been chosen:

L=180 microhenries.
Z=160,000 ohms (impedance of the tuned circuit).
F=1000 kilocycles (1,000,000 cycles).
Ra=8 ohms.

The Rm values that appear are the total impedances of the tube and the winding of the primary coil. Study the following table carefully and note the relation between the values which are given. These illustrate clearly just how the primary acts on the secondary.

Rm	$\omega^2 L^2(Z+Rm)$	Z+Rm	Ra+Rs
25,000	$.2367\times10^{12}$	$.004\times10^{12}$	59.20
50,000	$.269\times10^{12}$	$.008\times10^{12}$	33.60
500,000	$.845\times10^{12}$	$.08\times10^{12}$	10.56
1,500,000	2.125×10^{12}	$.24\times10^{12}$	8.85

The value of Rs which is in series with Ra may be found by subtracting the value of Ra (8 ohms) from the Rs+Ra value in the foregoing table as follows:

59.20−8 ohms=52.2
33.60−8 ohms=25.6
10.46−8 ohms= 2.56
8.85−8 ohms= .85

A curve showing the relation between (Ra+Rs) and Rm is shown in Fig. 4. Study this carefully and compare it with the foregoing figures.

With unity coupling (maximum or with a coefficient of 1) the AC plate resistance of the amplifying tube used is essentially the figure used for Rm. Of course, the primary reactance enters into this figure as well as the AC plate resistance. At resonance the secondary adds enough *capacity reactance to the primary* to neutralize the inductance so that it becomes

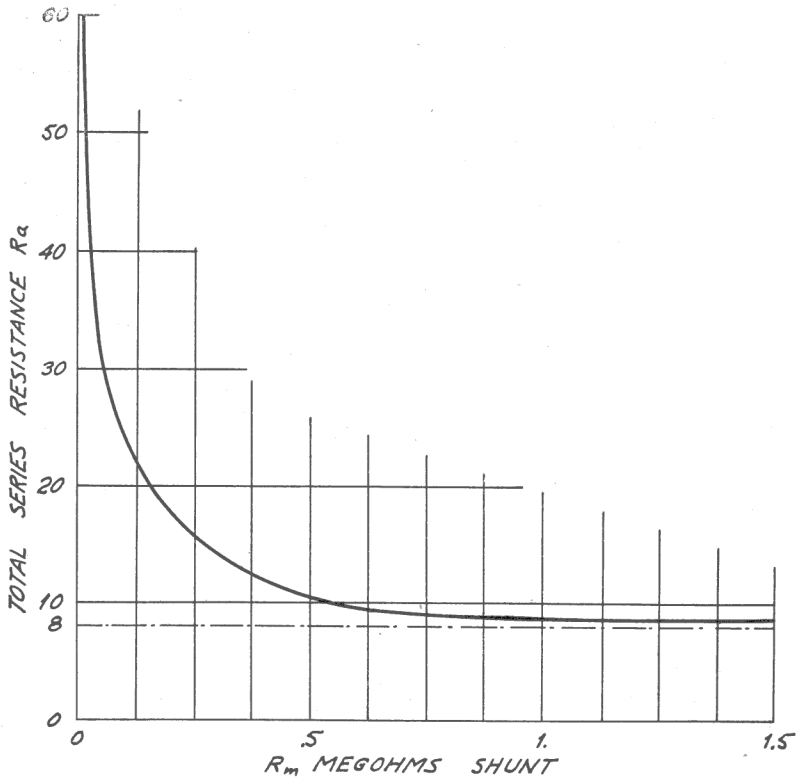


FIG 4

simply a resistance adding in series with the Rp of the tube. The primary, of course, affects the tuning of the secondary thru mutual inductance. The secondary must, therefore, be designed with consideration as to the mutual inductance and its effect on the secondary impedance.

The most basic thing to be learned from Fig. 4 and from the foregoing is the fact that the higher the RM value, the less the Rs value in the secondary circuit. This means that if a screen grid or pentode tube is used in the RF stage that its Rp will be high and, consequently, *the less series resistance there will be in the tuned secondary circuit.* It is, of course, very desirable to have as little resistance in the secondary circuit as possible. *Remember, however, that the action of the primary circuit does not change the Ra value—it affects only the Rs value.*

From all of the foregoing facts, it may be seen that everything about the secondary affects the primary and visa versa. You should not think of step-up or step-down ratio in RF coil design because of the variable mutual inductance and the phenomena of resonance. In fact, the volts-per-turn ratio in an iron core transformer would be destroyed if it were possible to acquire any appreciable resonance in either the primary or secondary circuit.

You are now ready to consider practical RF coupling systems. In Fig. 5 is shown one system wherein the tuned circuit is in series with the plate lead and the signal is fed to the following grid through the coupling condenser C1. This is known as a parallel resonant circuit. The grid cannot be left without a load because of accumulating electrons, and so a resistor R or grid leak is used in Fig. 5 to maintain a bias on the grid and to allow leakage of surplus electrons back to the cathode. The shunt resistance Rp of the first tube plus the shunt resistance of the combination C1−R, to say nothing of the additional load formed by the grid-cathode of the second tube, builds up a fairly large effective series resistance in the tuned circuit, lowering its shunt impedance and reducing its selectivity in addition to reducing the gain of the stage because of the decreased plate load.

One form of the series resonant circuit, as in Fig. 6, might be used if it were not for the large drop across R preventing the plate from being supplied with enough voltage for high gain. There is an advantage in the use of the tap on coil L which couples the grid circuit of the second tube to the plate of the first tube through C1. In this way, only part of the coil L is loaded with the Rp of the tube, allowing its total impedance to be higher which, of course, permits higher signal voltage on the following grid with better selectivity.

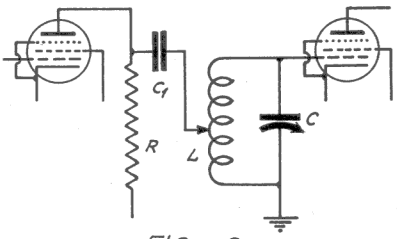


FIG. 6

In Fig. 7, many of the difficulties of Fig. 6 have been overcome. Full plate voltage may be applied to the plate of the first tube through the RF choke L1. Either a separate coil, as shown, or a tap may be used in this type of circuit. Fig 7 is another form of a series resonant circuit.

Obviously with changing frequencies and adjustment of the tuning condenser for these frequencies, there are going to be certain variations in gain of the circuit, due to changes in $\frac{L}{C}$ ratios, Q values, impedance variation, etc. Many methods have been suggested to overcome these difficulties. You know that as far as frequency is concerned, the reactance of a coil is opposite in nature to that of a condenser. The *large primary inductance* often used for screen grid and pentode RF amplifier

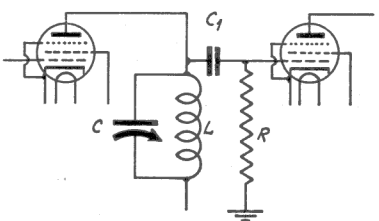


FIG. 5

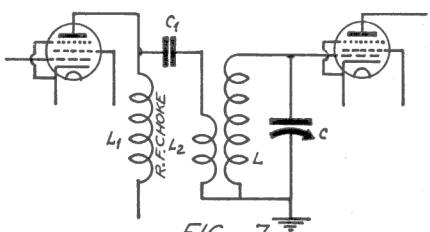


FIG. 7

tubes tends to loose its value at high frequencies, due to distributed capacity and near-resonant conditions produced at low frequencies. To improve the energy transfer at higher frequencies, capacitative coupling is employed with inductive coupling.

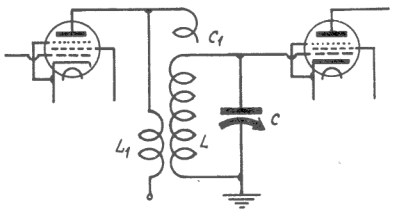


FIG. 8

This system is shown in Fig. 8, and you will remember you studied this form of coupling in a previous lesson. To permit the use of this form of coupling a small capacity is formed by winding a few turns of wire around the grid end of L in Fig. 8 to furnish capacitative coupling. This, of course, is more effective at high frequencies because of its reduced reactance and compensates for the decreasing effect of the primary in transferring energy to the secondary. In some cases all of the energy is transferred by capacity coupling of this sort—the primary not being coupled to the secondary at all. It is simply turned at right angles to the primary. Fairly even coupling characteristics are usually obtained in this way as the increased energy transfered at high frequencies is compensated out by the higher inherent circuit losses at high frequencies. There are, of course, numerous other workable circuits, but their cost or complexity makes them impractical although they may give somewhat superior results.

TUNED PRIMARY AND SECONDARY CIRCUITS

Having considered the controlling factors for good RF transformers and amplifier circuits with untuned primaries, you are now ready to consider transformers for RF and IF use, using both tuned primary and secondary windings.

Being determined by the relative amount of series resistance in a coil, its efficiency or general merit for use as a tuned coil may be expressed as $\frac{\omega L}{R}$. If it were possible to double the inductance without changing the resistance of the wire winding of a coil, the merit value of it would be doubled. This expression, $\frac{\omega L}{R}$, is called the Q of the coil as mentioned before, and, as practically all of the series resistance is in the coil, this expression may also be used as an appraisal of a complete tuned circuit.

To more thoroughly clarify this expression and its use, it may be evaluated with reference to any practical problem. A coil of 250 microhenries tuned with a 346 mmfd. condenser (tuned to 540 KC) will have an average series resistance of 4 ohms. Thus, to find the Q of the coil for these conditions, the solution is obtained as follows:

$$Q = \frac{\omega L}{Rs}$$

$$= \frac{6.28 \times F \times L}{Rs}$$

$$= \frac{6.28 \times 54 \times 10^{4} \times 25 \times 10^{-5}}{4}$$

$$= \frac{628\times54\times25\times10^{-3}}{4}$$

$$= \frac{33912\times25\times10^{-3}}{4}$$

$$= \frac{847{,}800\times10^{-3}}{4}$$

$$= \frac{847.8}{4} = 211.95$$

This value, 211.95 is the Q of the coil or tuned circuit. Obviously, the higher this Q value, the better will be the efficiency of the circuit. It will not only have a sharp resonance point, but its shunt impedance will be adequate for ordinary purposes.

Due to the flow of high frequency current and its effect on the series resistance, the Q of a coil will not remain constant over a large tuning range. It will vary over the broadcast band as much as 25 or 30%, averaging a value or around 120 for a well designed coil. Circuits can be built with Q values well above 800, but their size and expense does not justify their use in broadcast receivers. This may be done by using stranded wire, called Litz, after its inventor Litzendraht. From 7 to as many as 30 separately fine enameled wires form a single Litz conductor. Without increasing the ultimate bulk or cross section of the conductor, the *surface area is increased many times. This accounts for very much reduced radio frequency resistance* of the stranded wire even with a slightly higher DC resistance.

It is increasingly difficult to design a *high Q* tuned circuit at higher frequencies, as the RF resistance of the coil rises very rapidly with frequency as indicated

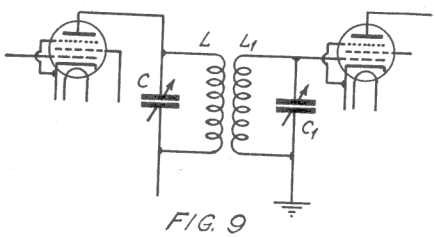


FIG. 9

by the curves in Fig. 2. A Q of 200 is considered good for the broadcast band while 75 to 100 is very good at 8 or 10 megacycles.

Concerning IF stages in super receivers refer to Fig. 9. As is the usual practice, assume that both tuned circuits are alike and are tuned to 460 KC. First consider them *separate or uncoupled.* For this case, values of C, L, C1 and L1 are taken from a standard IF coil.

C and C1=85–140 mmfd. adjusted to 110 mmfd.

L and L1=2,120 microhenries.

Frequency=460 KC (46×10^4).

RF coil resistance (Ra)=15 ohms.

$$Q = \frac{\omega L}{R} = 216$$

The shunt impedance of either circuit taken alone would be equal to:

$$Z = \frac{\omega^2 L^2}{Ra}$$

$$= \frac{39.4\times(46\times10^4)^2\times(112\times10^{-5})^2}{15}$$

$$= \frac{39.4\times2116\times10^8\times12{,}544\times10^{-10}}{15}$$

$$= \frac{39.4\times2116\times12{,}544\times10^{-2}}{15}$$

$$= \frac{83370.4\times125.44}{15}$$

$$= \frac{10457982.976}{15}$$

$$= 697{,}198+\text{ohms.}$$

If the tube to the left in Fig. 9

has an Rp of 800,000 ohms, it will affect the tuned circuit as though this much resistance were shunted across it. This happens because there is practically no impedance from B+ to the cathode of the tube and the top of the L-C circuit is connected to the plate. Now see how much the tube has increased the effective Ra of the circuit by means of the following:

$$Z = \frac{.697 \times .8}{.697 + .8} = \frac{.5576}{1.479}$$

$= .377$ megohms or 377,000 ohms.

For simplicity in working the foregoing problem for the two impedances in parallel, all values are expressed in *megohms*. Thus, if Z is now equal to 377,000 ohms the equivalent Ra may be found as follows:

$$Ra = \frac{\omega^2 L^2}{Z} = \frac{10,457,982}{377,000}$$

$= 27.74$ ohms.

The tuned circuit L1-C1 is exactly like L-C and hence alone would have a shunt impedance of 697,198 ohms at resonance. However, across this circuit is placed the grid-cathode circuit. The inherent grid to ground capacity through the cathode, screen and all other elements, shunts the tuned circuit and adds to the tuning capacity. The grid draws no current essentially and the impedance which it presents to the tuned circuit is around 2 megohms. This value will depend largely on the signal strength, as it does not remain constant for all signal intensities. With this as a reasonable value, there will be a series resistance in the circuit of—

$$Z = \frac{697,198 \times 800,000}{697,198 + 800,000}$$

$$= \frac{557758400000}{1497198}$$

$= 372,534+$ ohms

To find the resistance of Ra for this condition the solution is found as follows:

$$Ra = \frac{\omega^2 L^2}{Z}$$

$$= \frac{10,457,982}{372,534}$$

$= 28.01$ or 28 ohms.

The two tuned circuits of Fig. 9 have been purposely considered separated so that it will be perfectly clear to you what factors affect their design.

Next it may be assumed that the two tuned coils of Fig. 9 are brought within magnetic influence of each other. Start with .1 of 1% coupling or a coefficient of .001. This will create a mutual inductance of $\sqrt{L \times L1}$ which is equal to $1120 \times .001 = 1.12$ microhenries. (L and L1 being alike it makes $L = \sqrt{L \times L1}$ or $L1 = \sqrt{L \times L1}$.

The secondary will superimpose a resistance in series with the primary or it may be regarded in this way because it has an equivalent effect. Its value is found as follows:

$$Rs = \frac{\omega^2 L^2}{Ra}$$

$$= \frac{(6.28 \times 46 \times 10^4)^2 \times (1.12 \times 10^{-6})^2}{28}$$

and this is equal to:

$$\frac{39.4384 \times 2116 \times 10^8 \times 1.2544 \times 10^{-12}}{28}$$

$$= \frac{39.4384 \times 2116 \times 1.2544 \times 10^{-4}}{28}$$

$$= \frac{.394384 \times .12544 \times 211.6}{28}$$

$$= \frac{.049298 \times 211.6}{28}$$

$$= \frac{10.5314568}{28}$$

$= .3761$ ohms.

This broadens the tuning of the primary and, of course, the primary affects the secondary as though a resistor of the same value were placed in series with it. Because of the very small coupling, the amount of effective series resistance added is quite small.

Now, suppose that the coupling is increased between the two coils continuously until the resistance brought about in the primary, due to the presence of the secondary, amounts to as much as the resistance already there by reason of the tube plate resistance and the coil high frequency resistance. You have seen that the primary series resistance Ra is 27.74 ohms—15 ohms due to the circuit entirely isolated—the rest 12.74 ohms due to the Rp of the tube.

When the coupling is increased to a point where the secondary *couples into* the primary, a resistance equal to the resistance already there—*a maximum current will flow in the secondary and a maximum power will be transferred from one circuit to the other.* It is easy to mathematically find this amount of coupling, which is called the *critical coupling.* It is done simply by writing the expression for the quantity coupled into the primary from the secondary involving the mutual inductance, equal to the primary series resistance. If this then is solved for the mutual inductance, you can find the coupling coefficient by comparing the mutual inductance to the inductances of either L1 or L inasmuch as they are alike.

The procedure is as follows:

$\frac{\omega^2 M^2}{Rs} = Ra$ and transposing,

$$\omega M = \sqrt{RsRa}$$

Next solve for M as follows:

Rs=28 as worked out formerly.

Ra=27.74 at resonance.

$\omega = 6.28 \times 10^4 = 2{,}888{,}800.$

$$M = \frac{\sqrt{RsRa}}{\omega} = \frac{\sqrt{28 \times 27.74}}{2{,}888{,}800}$$

$$= \frac{\sqrt{776.72}}{2{,}888{,}800}$$

$$= \frac{28.8}{2{,}888{,}800}$$

$=.0000099$ or 9.9 microhenries.

It is obvious that the ratio $\frac{M}{L}$ will give the coefficient of coupling (K) where the two inductances are alike, and, in this case, it is equal to:

$$K = \frac{M}{L} = \frac{9.9}{1120} = .00803.$$

In spite of the fact that this is called *critical coupling,* it represents the maximum possible secondary current and power at resonance, yet the coupling is quite low. The resonance curves for current flowing in or voltage across the tuned circuits would appear as in Fig. 10.

What will happen if the coupling is increased beyond this point? When each coil or circuit couples into the other a resistance greater than what it already has, *the current at resonance will be reduced.* The secondary so loads the primary that the primary current is *reduced.* This, of course, prevents it from inducing as much en-

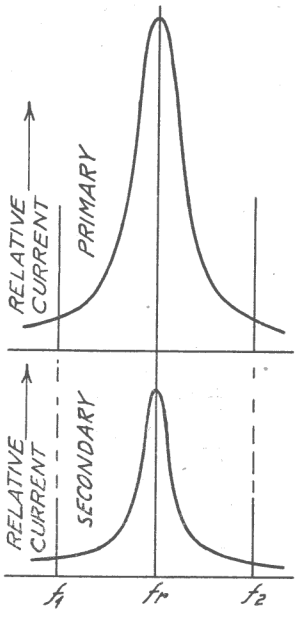


FIG. 10

ergy in the secondary and a low stable value is reached. However, as soon as the frequency is shifted slightly off resonance, there is no longer a *net resistance.* Instead of there being a net resistance coupling, there is a *net reactance* which couples into each circuit—the primary and secondary. The reactances adjust themselves to give a *net reactance which is either capacitive or inductive.*

At frequencies below resonance the series reactance of each circuit becomes capacitive, as this is *the nature of the net reactance.* In each, however, is coupled an inductive reactance in the form of M which couples the two and tends to balance the capacitive effects out of each circuit. The current is, therefore, *greater than at resonance,* and it would appear that the circuits have two resonant points as in Fig. 11. These, however, are not resonant points but characteristic wave band properties of the circuit when *over-coupled.* The current curves are shown in Fig. 11. Not only will you find the current is increased in both the primary and secondary for a point a trifle below resonance, but the same is true above resonance. In this case, the reactive influence in series with each circuit is predominately inductive, but the series impedance coupled into each by the other is *this time capacitive* in nature in spite of the fact that the mutual inductance is the transmitting medium for this capacitive reactance. Again, because of the mutual coupling, the C and L effects cancel one another as far as series action in the circuit is concerned.

If, however, the frequency is continuously decreased below point f1 in Fig. 11, the series capacitive reactance of each circuit rises more rapidly, and the primary current is too low to induce enough voltage in the secondary. Hence, the M does not support enough reactance to cancel the rapidly rising Xc. This rising series Xc soon reduces the primary and secondary currents to extremely low values. This happens only at frequencies a slight percent off resonance.

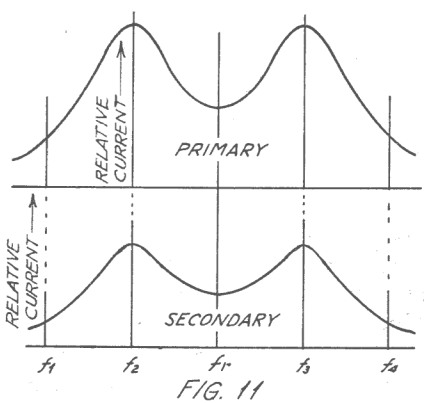


FIG. 11

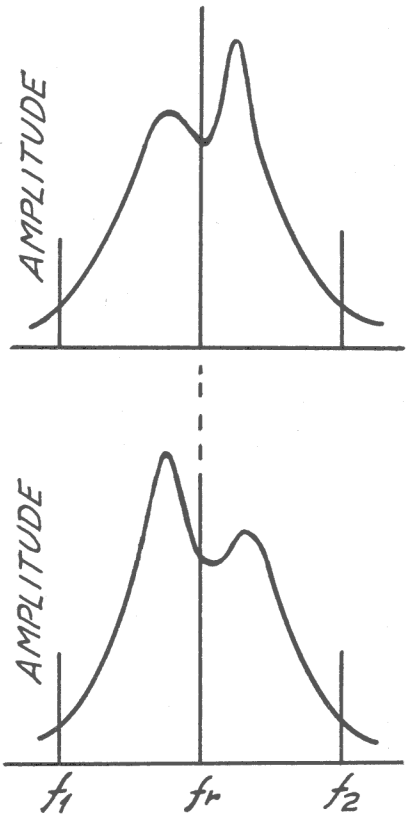


FIG. 12

In the upper frequency region, immediately above resonance, the same phenomena occurs as far as the current in the two circuits is concerned. This time the rapidly rising X_L above f2, in Fig. 11, cannot be counteracted by the reduced tendency for M to carry Xc into the circuits.

When coupling is increased from the critical value, the two *humps* appear in the resonance curve as in Fig. 11, and they spread apart as the coupling is increased—the resonant current steadily decreasing up to maximum coupling. When you start the study of the oscilloscope in your SAR Service Course, this subject of double hump curves will come up again—so be sure to remember what causes it.

Up to this point conditions have been considered only with both the primary and secondary tuned to resonance. If the coupling is at or below the critical value, the tuning of one of the circuits slightly off resonance will throw the other off of its natural resonant position, an amount depending on the coupling between the two. If the coupling is greater than the critical value to start with, the new resonant curve will be *double-humped* to start with, but will be distorted to an unsymmetrical shape by the detuning of one of the circuits, as in Fig. 12.

SELECTIVITY AND NATURE OF A MODULATED WAVE

In a previous lesson you learned that by connecting several tuned circuits in cascade in conjunction with amplifiers, it was possible to sharpen the selectivity almost indefinitely. How much selectivity is necessary and how may the right amount be obtained?

Early in the history of the transmission of intelligence by means of carrier waves for which were needed tuning mechanisms, it was discovered that the selectivity not only had an effect on the carrier but on the AF signal as well.

Thus, a carrier is modulated with a given fixed frequency, and then, if this carrier is amplified by means of a selective tube amplifier, the latter may affect the original modulation. It must be understood from the beginning of this part of the study that it deals with relative quantities rather than with

absolute values. For example, it is impossible to build a stable circuit which is infinitely sharp in tuning. One circuit may have greater selectivity than another.

The selectivity of a typical modern superheterodyne is indicated in Fig. 13. To digress for a moment, it will be recalled that if the sound level or sound power of any speaker, or, in fact, a sound wave from any source, decreases 1 db. a noticeable difference in sound level can be distinguished whereas if the change is less than 1 db (approximately 26% higher or lower), the change is not noticeable.

When a high frequency is modulated with a voice or other frequency for the transmission of some type of intelligence, the modulated wave requires a transmission *band* rather than a *knife edge* section of the total transmission spectrum. *Without modulation, the carrier wave may* be confined to circuits having no *lateral dimensions* or band width as it is called. They may be tuned to 1000 KC for a 1000 KC signal, for example, and may completely cut out the unmodulated RF signal if tuned to 1000.5 KC.

Such circuits would filter out a certain portion of almost all modulation placed upon them, because for usual transmission a band twice as wide in frequency as the highest frequency of modulation is required. For example, if a 1000 KC carrier is modulated with a 3000 cycle tone or a voice signal, it will require a transmission band of 6,000 cycles for proper transmission with the 1000 KC frequency at the center and extending 3000 cycles both above and below this carrier frequency. In other words, the higher frequency circuits must essentially amplify to the same degree and pass in every way all frequencies from 997 KC to 1003 KC. If the circuits will pass 10 times as much energy at a fixed carrier frequency of 1000 KC as at 997 KC, it will not be satisfactory for a transmission of a 1000 KC carrier wave modulated by a 3000 cycle frequency. This is true not only for a signal of 3000 cycles but for all other modulating frequencies as well. The carrier frequencies of broadcast stations are spaced 10 KC apart to allow as many stations as possible within the section of the spectrum most useful for broadcasting and allow enough spacing between station carriers so that each station may have a *transmission band* or *channel* in which to transmit modulation. In addition to this, where the receiv-

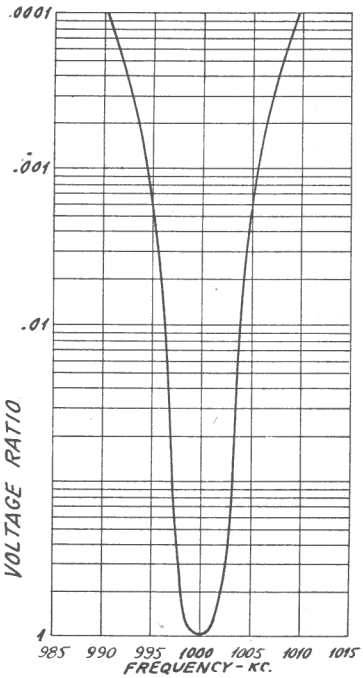
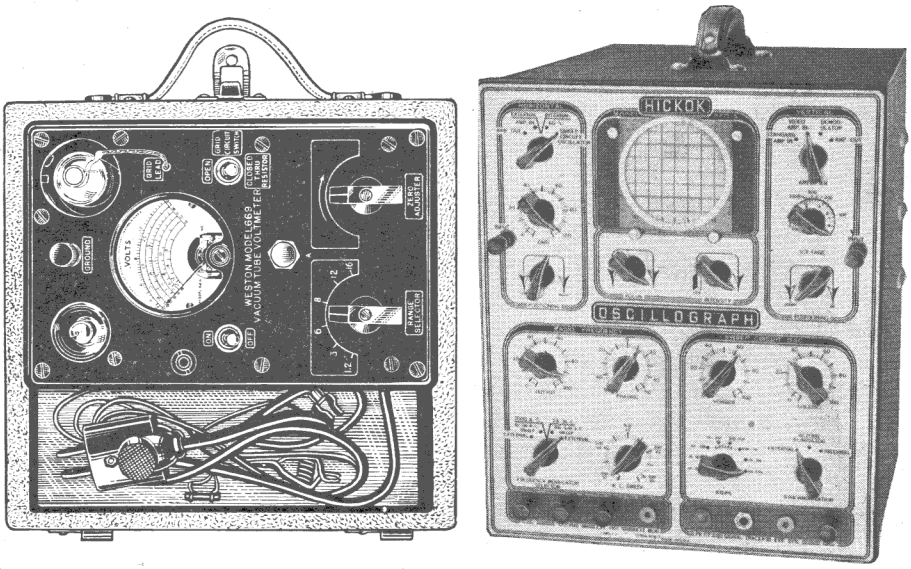


FIG. 13



Above are shown two test instruments that are employed frequently in adjusting the type of tuned circuit described in this lesson. The instrument on the left is a Weston vacuum tube voltmeter while the one on the right is a Hickok oscillograph. Full instructions for using these instruments are given in your SAR Service Course Lessons.

er has poor selectivity and is capable of picking up two adjacent waves, they will form an almost inaudible beat (10 KC). Of course, many people can hear this frequency but radio station frequency allocations are made in such a way that this type of interference is at a minimum.

Now it may be quite readily understood that if one station on 1000 KC is transmitting groups of modulation frequencies up to 5000 cycles and another one at 990 KC is transmitting modulation frequencies of 5000 cycles also, one would extend from 985 to 995 KC and the other from 995 to 1005 KC. They would, therefore, be directly against each other. If either used a greater modulation frequency, they would overlap and cause interference.

The reason why a modulated wave requires a transmission band is somewhat controversial, but the facts are, of course, susceptible to laboratory proof. There is no possible way to make any one reason valid or void by means of mathematical analysis as the result will only be an evaluation of the suppositions with which the problem was begun. You may start with any one of several suppositions in this case.

Instead of entering into this controversy or attempting to show one phase of this argument by means of complicated mathematical analysis, it is possible to simply take the facts as they come from the laboratory and piece them together as logically as possible. If an attempt is made to pass a modulated carrier through a highly selective circuit, it will be noticed that a definite change in the character of the modulation has taken place. If you have good ears and listen carefully, you may note further that the highest frequencies in the

group being transmitted have been attenuated or reduced. The extent of reduction of the higher frequencies, of course, is a matter of degree. A circuit cannot be selective enough to completely cut out *all* of any frequency, and so the amount of attenuation is related to the selectivity of the circuit and the value of the modulation.

The attenuation of the various frequencies of modulation of a carrier for a highly selective circuit would be somewhat like Fig. 14. They would be cut down in proportion to their relative values. Thus, the high modulation frequencies (3000 to 5000 cycles) are cut down considerably more than the low frequencies (40 to 3000 cycles). No matter how high the modulation frequency, there will always be some energy carried by the highly selective circuit, just as there will be some energy transmitted *off resonance* no matter how selective the circuit through which it is passed.

What are the characteristics of a circuit which will make it selective to the modulation of the carrier to which it is tuned? Figure 15 has been prepared to convey the nature of a highly efficient tuned circuit or group of tuned circuits. Due to the overall low resistance of any highly selective system, it has, for one thing, the nature of *fly-wheel* action or *self-inertia.* Once it has been started or energy has been placed in it, it tends to hold it indefinitely. This might be compared to the swing of a pendulum in air and water. There is so much friction or resistance to the motion of the pendulum under water that it may swing only a few strokes before it comes to rest. In the air, it may swing a hundred or so strokes before coming to rest. Thus, in an efficient circuit with the lowest possible series resistance in the tuned elements, electrical energy once started will oscillate for a relatively long time before coming to rest.

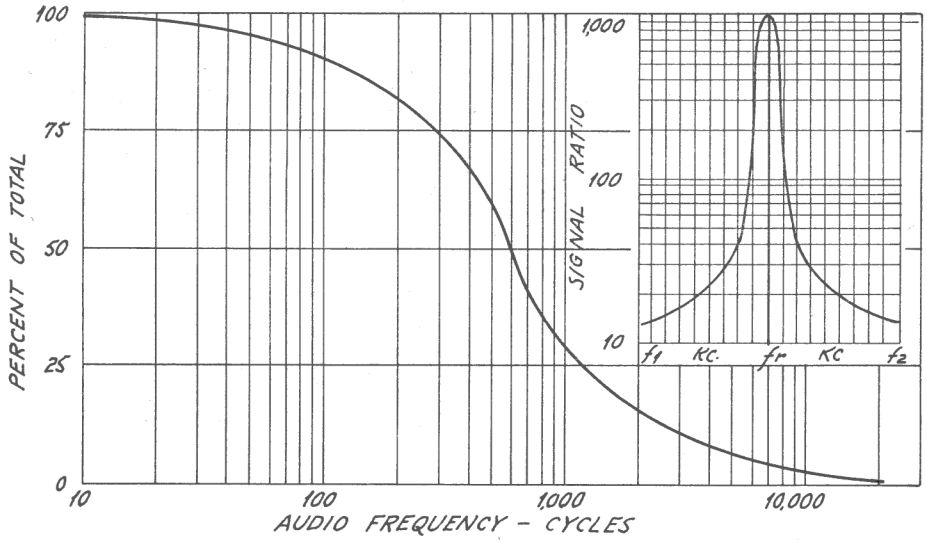


FIG. 14

According to your understanding of a modulated wave, to carry this energy a circuit must be able to *get rid* of its *self energy* as fast or faster than the supplied energy reduces. For example, use a 1000 KC wave modulated at 5000 cycles. The energy must build up to maximum in 1/10,000 of a second and die down to minimum in an equal time. The circuit must be such that any energy supplied to it will die down to zero in 1/10,000 of a second. If it requires a longer time, such as 1/2,000 of a second, when the modulation energy fed to it reduces to zero or minimum, there is still some energy in the circuit due to the self-inertia of the circuit. If a resistance is placed in the circuit in any way, it will *quench or stop self-inertia.* In this way, the energy may rise and drop as rapidly as the energy fed to it rises and drops.

In this way the circuit may be made to carry higher modulation frequencies, depending on the amount of resistance it has. This, of course, checks perfectly with the fact that a circuit having considerable series resistance tunes broadly. Over an almost infinitely short time the *residual* energy of any such tuned system will drop a certain amount and hence can carry some energy of almost any high value modulation frequency. However, it will be seriously attenuated or cut down.

A tuned system opposes any change in the amount of high frequency energy which it contains just an inductance opposes any change in current flow through it. Modulation has the effect of changing the amount of energy in the wave, and the more rapidly it is changed, the more becomes the opposition of the tuned circuit to transmit the variations. Thus the ratio of the amplitude of the modulation to the carrier energy is decreased as it becomes higher in frequency just as an inductance will have increasingly higher reactance at high frequencies.

Now look at this same situation from a strictly mathematical view point. It has been covered in the foregoing in a strictly physical or mechanical way. Equations show

An efficient antenna system will always bring in more signal energy free from noise than will an inefficient system. On the left in the above view is shown an antenna coupling transformer—used to couple the antenna system to the input of the receiver—to match impedances. On the right a complete antenna system is shown with matching transformers, transmission line, etc. Courtesy of Technical Appliance Corp.

that the sum of two waves, the carrier and the modulating wave will produce two additional waves—one equal to their sum and one equal to their difference. As you will remember, this is exactly what was studied in connection with the superheterodyne except that then the problem was dealing with two waves of high frequency. At that time it was mentioned that any two frequencies could be treated in this sense. With the present case there are two frequencies of 1000 KC and 5 KC being mixed or heterodyned, forming two additional frequencies of 1005 KC and 995 KC which are the sum and difference frequencies.

In connection with modulated waves, these two are called *side bands* because they are equally spaced either side of the carrier wave. Either one or both of these side bands may be amplified and used to transmit the modulated intelligence to a detector system. In connection with your superheterodyne studies, only one side band was used such as 175 KC (the IF frequency). This and another frequency of 2175 KC may be formed by heterodyning a 1000 and 1175 KC frequencies. The *upper side-band* (2175 KC) is ignored while the *lower side-band* (175 KC) is used for the IF as explained.

In the case of a modulated signal it is desired to transmit each side-band and all of the other side-bands which might be produced between 40 cycles and 5000 cycles. For this reason the tuning circuits must transmit a maximum band of frequencies 10 KC wide.

Ten KC is only 1% of 1000 KC which represents a very high selectivity at 1000 KC, but it is 10/175 or 5.72% of 175 KC and thus more easily attained in a 175 KC circuit than in a 1000 KC circuit. By reducing the frequency used as a carrier of the modulation, it is possible to gain quite an advantage in selectivity.

The problem now becomes one of obtaining *as high a selectivity as possible in order to properly select stations but broad enough to carry all of the intelligence being modulated on the carrier.* Ordinarily if the selectivity of a circuit is made fairly great, it will have a shape which will *make it favor low frequencies more and more.* For this reason, at resonance the selectivity should be as broad as possible so that the resonant conditions will extend out to 5000 cycles in order to cover each side-band. The shape of the resonance response curve must have a flat top section or as nearly so as possible so that the amplifier may not *seleceively attenuate* or amplify any frequencies in the audio spectrum.

One way of doing this is to stagger the various IF tuned stages as in Fig. 15 so that one will have a peak, for example, 4 KC below resonance, one on resonance and one 4 KC above resonance. The total selectivity will be as the solid line in Fig. 15 indicates. This results in reduced gain because each circuit must carry the entire wave, and although flat response, as this is called, will be acquired, it will be at the expense of reduced gain. This staggering may be done in the stages of the IF amplifier. It will not be beneficial in the RF tuned system.

As you have just learned, overcoupling (staggering the tuned

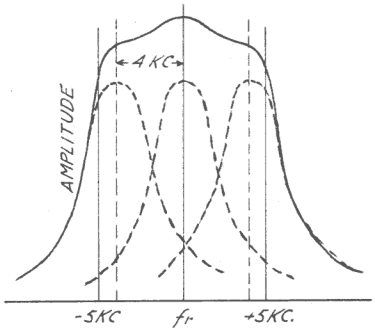
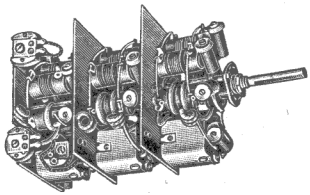


FIG. 15

stages) will depress the resonant point but not other points materially and is a good system to use. Still other methods involve the use of resistance actually placed in series with one or more tuned circuits to broaden their response characteristics. Metalic cores of *permalloy* or other trade names, as in Fig. 16, are often used to add resistance to tuned circuits at resonance. Off resonance the fields are reduced to such an extent that the core material comes into action. The same is true of a copper disc placed between the two windings of an IF transformer as in Fig. 17. At resonance it picks up considerable power from the tuned circuits as eddy currents flow in the copper disc due to induced voltages. The eddy current field forms mutual inductance which adds to the resistance of the tuned circuit at and near resonance. Iron cores in different forms are used with IF transformers for the same reason as the copper disc. Two of these iron core arrangements are also shown in Fig. 17.

For high fidelity transmission covering as high a band as 30 KC (15 KC each side of the carrier) a third winding is used, as in Fig. 18, just like the primary and secondary and coupled to both. It is tuned but has a resistance in series with it usually adjustable to the most desirable value. From 8 to 10 ohms is sufficient for this purpose. At resonance it dissipates much more energy than at any other frequency. Slightly off resonance it has practically no effect on either the primary or secondary of the circuit. It serves to give the tuned stages a *flat top* characteristic in the response curve, as in Fig. 15, which is highly desirable for high fidelity reproduction.



The above view shows a compact Messiner tuning system. Such a unit covers several bands of frequencies. You will learn more about these when when you take up the study of all wave receivers.

NUMBER OF TUNED STAGES AND GAIN OF IF AMPLIFIER

Some of the very small superheterodyne circuits have no intermediate frequency amplifier at all. The plate circuit of the mixer or first detector is tuned, and this is coupled to a tuned input to the second detector. Thus, there is no gain in the IF stages. The second

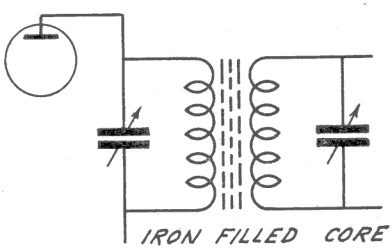


FIG. 16

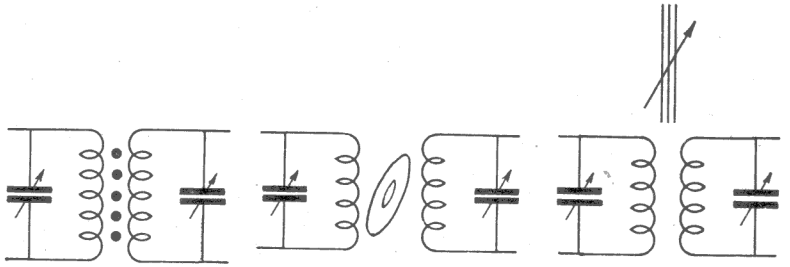


FIG. 17

detector is, in this case, of the power type, having some gain as well as detector action. A diode would not be suitable here because it would have no gain. As many as three IF stages are used in other larger superheterodynes. These are for all-wave super-sensitive reception. Two IF stages with a gain of 75 to 90 each, making a total gain of 5600 to 8100, are adequate for all ordinary purposes.

It is rather difficult for the average student on studying this lesson for the first time to appreciate the full significance of the principles of this lesson. Thus, it will be well worth your time to study it again and again.

One of the most valuable functions of this lesson is to serve as a reference source. Whenever a problem of tuning, resonance or coupled circuits comes up, be sure to refer back to this lesson. By doing so you can, no doubt, find an answer to your problem.

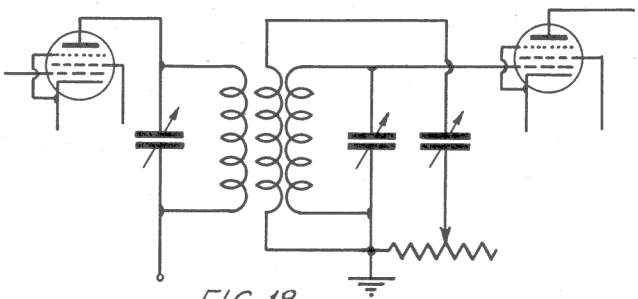


FIG. 18

These questions are designed to test your knowledge of this lesson. Read them over first to see if you can answer them. If you feel confident that you can, then write out your answers, numbering them to correspond to the questions. If you are not confident that you can answer the questions, re-study the lesson one or more times before writing out your answers. Be sure to answer every question, for if you fail to answer a question, it will reduce your grade on this lesson. When all questions have been answered, mail them to us for grading.

QUESTIONS

No. 1. What is the main factor in a tuned circuit which governs the ratio of inductance to capacity for the highest possible efficiency?

No. 2. In what part of a tuned circuit is the most resistance to be found and what causes it?

No. 3. What kind of wire or conductor should be chosen for a coil for it to have minimum radio frequency resistance?

No. 4. How will a resistance in parallel with a tuned circuit affect the current in that circuit?

No. 5. What is meant by the "Q" of a tuned circuit?

No. 6. What is the ultimate effect of plate circuits, grid circuits and coupled coils on a tuned circuit with respect to the series resistance of the circuit?

No. 7. If two circuits tuned to the same frequency are over coupled, will the maximum current flow at resonance?

No. 8. Can the combined selectivity of a group of circuits have any effect on the audio modulation of the carrier? Explain.

No. 9. What is meant by critical coupling of two tuned circuits?

No. 10. Name three ways of broadening the frequency response of tuned circuits to prevent attenuation or reduction of the modulation frequencies.